Turbo Coded Space-time Block codes for four transmit antennas with linear precoding

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***Abstract*— In this paper, we combine Turbo Codes (TC) and Space-Time Block Coding (STBC) in order to exploit channel coding and spatio-temporal diversities. Moreover, linear precoding matrices are combined to efficiently improve the space-time diversity order of the multiple antenna system. In addition of these good performance results, the proposed system implements a simple linear decoder at the reception side, which has very low complexity compared to the Maximum Likelihood (ML) detectors currently used in such systems. Performance results, obtained by simulations are given for a 4 transmit antenna system, with different precoding matrix sizes over flat Rayleigh fading channels, for a global system including turbo channel coding. The described scheme can easily and efficiently be adapted to a two transmit antenna system. The results are compared to other schemes given in literature.**

I. Introduction

Since 1993, Turbo codes are demonstrated to be very efficient to exploit the coding diversity, performing close to the Shannon limit [1]. In 1996, Foschini demonstrated that the spectral efficiency increases linearly with the minimum of transmit or receive antennas [2] when using multiple antenna systems. Another way to exploit multiple antennas is to use STBC.

The initial two-transmit antenna Orthogonal STBC (OSTBC) proposed by Alamouti [3] is merely decoded with a linear operation. Another advantage of this code is its unitary rate. Then, Tarokh [4] extended OSTBC to 3 or 4 transmit antennas with linear decoding as well, but resulting in lower 1/2 and 3/4 rate codes. Since, many studies have been carried out to find out space-time codes with more than two antennas and rate one, all resulting in Quasi Orthogonal STBC (QOSTBC) [5, 6, 7], thus requiring more complex than linear detector such as Maximum Likelihood (ML) detector.

In parallel, linear precoding also called constellation rotation was demonstrated to efficiently exploit time diversity for Single Input Single Output (SISO) systems [8]. In [9], a system combining the STBC proposed in [5] with linear precoding is presented, but the complexity of its ML detector varies exponentially with the length of the precoding matrix.

In this paper, we tested the influence of the channel coding for the system pesented in [10], where linear precoding and OSTBC are efficiently combined. As channel coding scheme, we have implemented the robust duo binary turbo channel encoder that is proposed in DVB-RCT standard [11]. This scheme is compared to linear precoded QOSTBC described in [9]. As we will show, the diversity order increases with the size of the precoding matrix. We will focus our attention on linear precoding matrices based on Hadamard matrix construction. This specific combination of turbo coded linear precoding and OSTBC has the effect of increasing the overall diversity of the system by scattering the information in the space, time or frequency domains. At the receiver, before turbo decoding a simple linear detector offers a good trade-off between performance and complexity thanks to low interference terms.

First, in session II, we recall the channel representation of OSTBC. In session III, we focus on the state of the art of 4 transmit existing systems and especially on QOSTBC channel representation. In session IV, we introduce our new scheme for 4 antennas, which is based on the combination of one particular linear precoding technique with OSTBC. Formulas are given, highlighting the importance of the interference terms that must be minimized in order to increase the performance. Before concluding, we give simulation results of the proposed system for different precoding matrix sizes, with and without duo binary turbo channel coding. We compare the obtained performance to existing four antennas systems.

For this study, frequency non-selective flat Rayleigh fading channel and time invariance during T symbol durations are assumed as well as perfect channel estimation. Hence, the theoretical channel response, from transmit antenna t to receive antenna r can be estimated by $h_{tr} = \rho_{tr} e^{i\theta_{tr}}$. Moreover, we consider uncorrelated channels from each transmit antenna t to each receive antenna r.

II. General OSTBC description

In this section an OSTBC description based on channel representation is given.

The Alamouti code [3] can be represented as follows:

$$\mathcal{G}_2 = \begin{bmatrix} s_1 & -s_2^* \\ s_2 & s_1^* \end{bmatrix} \quad (1)$$

Under the assumption that the fading coefficients are constant over $T = 2$ consecutive symbol durations, the Alamouti OSTBC code for $N_t = 2$ transmit and $N_r = 1$ receive antenna is represented by:

$$\mathbf{r} = \mathbf{H}.\mathbf{s} + \mathbf{n} \quad (2)$$

where $\mathbf{r} = [r_1\, r_2]^T$ is the received signal over two consecutive symbol durations, $\mathbf{s} = [s_1\ s_2]^T$ is the transmitted signal, $\mathbf{n} = [n_1\ n_2]^T$ is the additive white gaussian noise,

$$\mathbf{H} = \begin{bmatrix} h_1 & h_2 \\ -h_2^* & h_1^* \end{bmatrix} \quad (3)$$

is the equivalent channel matrix for the 2 successive symbol durations over 2 antennas, and h_i is the channel response of the transmit antenna i.

The decoding step consists in applying the transpose conjugate of the channel matrix to the equivalent received vector. The Alamouti decoding is performed by:

$$\hat{\mathbf{s}} = \mathbf{\Lambda}.\mathbf{s} + \mathbf{n}' \quad (4)$$

where $\hat{\mathbf{s}} = [\hat{s}_1\ \hat{s}_2]^T$ is the estimated symbol vector after decoding, $\mathbf{\Lambda} = \mathbf{H}^H.\mathbf{H} = \lambda.\mathbf{I}_2$, where $(.)^H$ stands for the transconjugate, $\mathbf{I}_2$ the identity 2x2 matrix, $\lambda = |h_1|^2 + |h_2|^2$ and $\mathbf{n}' = \mathbf{H}^H.\mathbf{n}$. In this paper, we provide formulas with the equalization process corresponding to a Maximum Ratio

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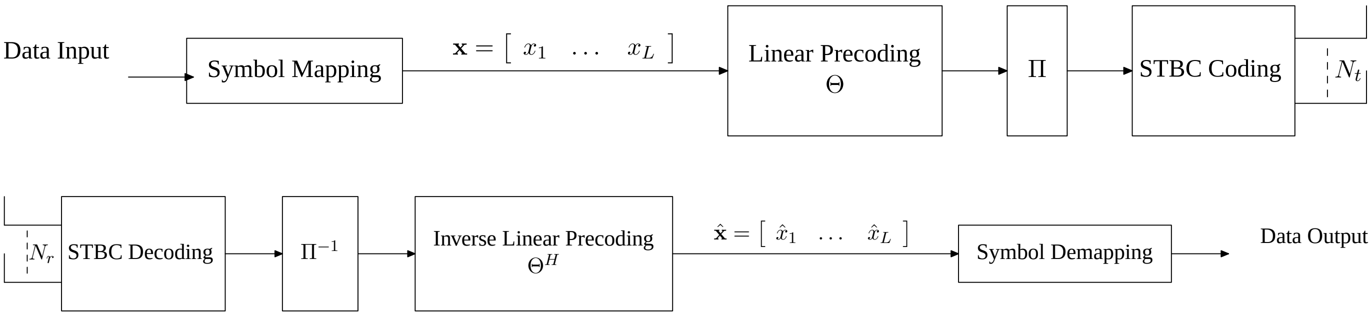

Fig. 1. Linear precoding with OSTBC transmitter and receiver

Combining (MRC) equalizer. However, an equalization process can be carried out according to the Zero Forcing (ZF) or Minimum Mean Square error (MMSE) criteria leading to different performance when using linear precoding.

III. Existing systems with Four-transmit antennas

STBC codes of rate 1 for 4-transmit antenna systems are not orthogonal. They can be expressed by (2) where $\mathbf{r} = [r_1\, r_2\, r_3\, r_4]^T$ is the received signal and $\mathbf{s} = [s_1\, s_2\, s_3\, s_4]^T$ the transmitted one. The total transmit power is P. Each antenna transmits a symbol over one symbol duration, therefore each antenna transmits symbol at a power of $P/4$. The matrix representation described in section II can be extended to these QOSTBC schemes [5, 6, 7]. For the QOSTBC proposed in [5] by Jafarkhani, $\mathbf{H}$ is therefore a 4x4 matrix equals to:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 \\ -\mathbf{H}_2^* & \mathbf{H}_1^* \end{bmatrix} \tag{5}$$

while the NOSTBC proposed in [6] by Tirkkonen is:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & \mathbf{H}_2 \\ \mathbf{H}_2 & \mathbf{H}_1 \end{bmatrix} \tag{6}$$

The equivalent channel matrix for 2 successive symbol durations over 2 antennas is:

$$\mathbf{H}_i = \begin{bmatrix} h_{2i-1} & h_{2i} \\ -h_{2i}^* & h_{2i-1}^* \end{bmatrix} \tag{7}$$

where h_{2i} and h_{2i-1} are the channel responses of transmit antenna $2i$ and $2i-1$ respectively, $i \in \mathbf{N}^*$.

In [9], the authors propose to apply a linear precoding to non orthogonal matrix [5] and to use a ML detector at the receiver part. The implementation complexity increases exponentially with the size of the precoding matrix. To sum up, all these OSTBC require ML or similar detectors due to high interference terms.

IV. Proposed scheme for four transmit-antennas

The system, combining the Linear Precoding with OSTBC (transmission and reception sides), is presented in Figure 1, without channel coding. When duo-binary turbo channel coding is implementing, the encoder is inserted before symbol mapping whereas the channel decoder is added after the symbol demapping module.

The global transmission/reception scheme is given by:

$$\mathbf{A}_L = \mathbf{\Theta}_L.\mathbf{\Lambda}_L.\mathbf{\Theta}_L^H \tag{8}$$

According to the theorem of diagonal decomposition, $\mathbf{A}_L$ is an Hermitian matrix and $\mathbf{\Theta}_L$ an unitary matrix used in the system as linear precoding matrix. When these unitary matrices are constructed according to the orthonormal complex Hadamard matrices, the diagonal terms of $\mathbf{A}_L$ are equal to the sum of the λ_i's, providing full diversity on the detected symbols. In addition, we remark that the interference terms correspond to the difference between these λ_i's.

A. Linear precoding based on Hadamard construction

We propose to use the following linear precoding based on the Hadamard construction matrix such as:

$$\mathbf{\Theta}_L = \sqrt{\frac{2}{L}} \begin{bmatrix} \mathbf{\Theta}_{L/2} & \mathbf{\Theta}_{L/2} \\ \mathbf{\Theta}_{L/2} & -\mathbf{\Theta}_{L/2} \end{bmatrix} \tag{9}$$

with $L = 2^n$, n $\in \mathbf{N}^*$, n ≥ 2 and:

$$\mathbf{\Theta}_2 = \begin{bmatrix} e^{j\theta_1}.\cos\eta & e^{j\theta_2}.\sin\eta \\ -e^{-j\theta_2}.\sin\eta & e^{-j\theta_1}.\cos\eta \end{bmatrix} \tag{10}$$

belonging to the Special Unitary group SU(2), therefore $det(\mathbf{\Theta_2}) = 1$. This leads to the following general expression:

$$\mathbf{A}_L = \frac{2}{L} \begin{bmatrix} \mathbf{A}_{L/2}^1 + \mathbf{A}_{L/2}^2 & \mathbf{A}_{L/2}^1 - \mathbf{A}_{L/2}^2 \\ \mathbf{A}_{L/2}^1 - \mathbf{A}_{L/2}^2 & \mathbf{A}_{L/2}^1 + \mathbf{A}_{L/2}^2 \end{bmatrix} \tag{11}$$

with

$$\begin{aligned} \mathbf{A}_{L/2}^1 &= \mathbf{\Theta}_{L/2}.\mathbf{\Lambda}_{L/2}^1.\mathbf{\Theta}_{L/2}^H \\ \mathbf{A}_{L/2}^2 &= \mathbf{\Theta}_{L/2}.\mathbf{\Lambda}_{L/2}^2.\mathbf{\Theta}_{L/2}^H \end{aligned} \tag{12}$$

$$\mathbf{\Lambda}_L = \begin{bmatrix} \mathbf{\Lambda}_{L/2}^1 & 0 \\ 0 & \mathbf{\Lambda}_{L/2}^2 \end{bmatrix} \tag{13}$$

where

$$\begin{aligned} \mathbf{\Lambda}_{L/2}^1 &= diag(\lambda_1, \ldots, \lambda_{L/2}) \\ \mathbf{\Lambda}_{L/2}^2 &= diag(\lambda_{L/2+1}, \ldots, \lambda_L) \end{aligned} \tag{14}$$

in the case of interleaving. Indeed, the interleaving has the effect of mixing eigenvalues between different blocks, thus the components of the resulting eigenvalue matrix are different from each others.

Therefore, for $L = 2$, we obtain the following hermitian matrix:

$$\mathbf{A}_2 = \begin{bmatrix} a & b \\ c & d \end{bmatrix} \quad (15)$$

with

$$\begin{array}{l} a = \cos^2\eta.\lambda_1 + \sin^2\eta.\lambda_2 \\ b = -\cos\eta.\sin\eta.e^{j(\theta_1+\theta_2)}.(\lambda_1 - \lambda_2) \\ c = -\cos\eta.\sin\eta.e^{-j(\theta_1+\theta_2)}.(\lambda_1 - \lambda_2) \\ d = \sin^2\eta.\lambda_1 + \cos^2\eta.\lambda_2 \end{array} \quad (16)$$

Since $\mathbf{\Theta}_L$ is an unitary matrix, the diagonal elements a_{ii} of $\mathbf{A}_L$ are similar to:

$$a_{ii} = \frac{2}{L}\sum_{l=0}^{L/2-1}\cos^2\eta.\lambda_{2l+1} + \sin^2\eta.\lambda_{2l+2}\, \forall i \in [1 \dots L] \quad (17)$$

while the interference terms are all difference between these λ_l's. One of these interference terms is:

$$a_{ik} = -\tfrac{2}{L}\cos\eta.\sin\eta.e^{-j(\theta_1+\theta_2)}\sum_{l=0}^{L/2-1}(\lambda_{2l+1} - \lambda_{2l+2}) \quad i,k \in [1 \dots L]_{i\neq k} \quad (18)$$

Owing to equation (13), the other terms of interference are also sum of differences between eigenvalues.

By simulation, we found that the optimal BER performance results were obtained with pure real or pure imaginary interference, leading to the chosen values $\eta = \frac{\pi}{4}$, $\theta_2 = \theta_1 - \frac{\pi}{2}$, $\theta_1 = \frac{5\pi}{4}$ giving the following matrix:

$$\mathbf{A}_2 = \frac{1}{2}\begin{bmatrix} \lambda_1 + \lambda_2 & \lambda_2 - \lambda_1 \\ \lambda_2 - \lambda_1 & \lambda_1 + \lambda_2 \end{bmatrix} \quad (19)$$

Therefore, with these coefficients, we get the following global formula:

$$\mathbf{A}_L = \frac{1}{L}\sum_{i=1}^{L}\lambda_i.\mathbf{I}_L + \mathbf{J} \quad (20)$$

with $\mathbf{I}_L$ the identity matrix of size $L \times L$ and $\mathbf{J}$ the matrix of interference terms.

One can see that diagonal elements are the sum of the eigenvalues and the interference terms are sum of difference of eigenvalues. This particularity is conserved when L increases. Moreover, for flat independent Rayleigh channels, when L increases the diagonal terms tend to a non-centered Gaussian law while the interference terms tend to a centered Gaussian law.

B. System description

In our 4-transmit antenna system, we first apply Alamouti OSTBC in order to keep the rate of our proposed scheme to one. The Alamouti OSTBC is applied alternatively to antennas 1 and 2 and then to antennas 3 and 4. Thus, the symbols are transmitted over the first group of transmit antennas 1 and 2, with a power $P/2$ over two symbol durations, when the other antennas are switched off. Then, the other symbols are transmitted over the second group of transmit antennas 3 and 4. Therefore, as for the previous QOSTBC codes with 4 antennas, the total transmit power per symbol duration is equal to P.

The corresponding OSTBC matrix representation for our one-rate four antenna system with a diversity order of 2 is:

$$\mathcal{G}_2 = \begin{bmatrix} s_1 & -s_2^* & 0 & 0 \\ s_2 & s_1^* & 0 & 0 \\ 0 & 0 & s_3 & -s_4^* \\ 0 & 0 & s_4 & s_3^* \end{bmatrix} \quad (21)$$

The channel representation of the OSTBC codes gives the equivalent channel coding matrix $\mathbf{H}$ of size 4×4:

$$\mathbf{H} = \begin{bmatrix} \mathbf{H}_1 & 0 \\ 0 & \mathbf{H}_2 \end{bmatrix} \quad (22)$$

At the reception side, in (4) we get $\mathbf{\Lambda} = \mathbf{\Lambda}_4 = \mathbf{H}^H.\mathbf{H} = diag(\lambda_1, \lambda_1, \lambda_2, \lambda_2)$ where $\lambda_1 = |h_1|^2 + |h_2|^2$ and $\lambda_2 = |h_3|^2 + |h_4|^2$ leading to a 2 channel diversity order. Before this space-time code, we apply a linear precoding represented by an $L \times L$ unitary matrix described in the previous section.

For $L = 4$ when applying the optimal coefficients, we obtain:

$$\mathbf{\Theta}_2 = \frac{1}{2}\begin{bmatrix} -1-j & -1+j \\ 1+j & -1+j \end{bmatrix} \quad (23)$$

and the global transmission and reception system is described by:

$$\mathbf{A}_4 = \mathbf{\Theta}_4.\mathbf{\Lambda}_4.\mathbf{\Theta}_4^H \quad (24)$$

$$\mathbf{A}_4 = \frac{1}{2}\begin{bmatrix} \lambda_1+\lambda_2 & 0 & \lambda_1-\lambda_2 & 0 \\ 0 & \lambda_1+\lambda_2 & 0 & \lambda_1-\lambda_2 \\ \lambda_1-\lambda_2 & 0 & \lambda_1+\lambda_2 & 0 \\ 0 & \lambda_1-\lambda_2 & 0 & \lambda_1+\lambda_2 \end{bmatrix} \quad (25)$$

where the diagonal elements of $\mathbf{A}_4$ are all equal to:

$$a_{ii} = \tfrac{1}{2}\sum_{l=1}^{L=4}|h_l|^2 \quad \forall i \in [1 \dots 4] \quad (26)$$

Therefore, owing to linear precoding, the exploited channel diversity order increases from 2 to 4. Moreover, interference terms are either null or similar to:

$$a_{ik} = \tfrac{1}{2}(\sum_{l=1}^{2}|h_l|^2 - \sum_{l=3}^{4}|h_l|^2) \quad i,k \in [1 \dots L]_{i\neq k} \quad (27)$$

If we take $L = 4$ with the Alamouti scheme and interleaving we obtain the following diagonal channel matrix:

$$\mathbf{A}_4 = \mathbf{\Theta}_4.\mathbf{\Lambda}_4.\mathbf{\Theta}_4^H \quad (28)$$

with $\mathbf{\Lambda}_4 = diag(\lambda_1, \lambda_2, \lambda_3, \lambda_4)$.

Then, the resulting matrix becomes:

$$\mathbf{A}_4 = \frac{1}{4}\begin{bmatrix} a+b & c+d & a-b & c-d \\ -c-d & a+b & -c+d & a-b \\ a-b & c-d & a+b & c+d \\ -c+d & a-b & -c-d & a+b \end{bmatrix} \quad (29)$$

with

$$\begin{array}{ll} a = \lambda_1 + \lambda_2 & b = \lambda_3 + \lambda_4 \\ c = \lambda_1 - \lambda_2 & d = \lambda_3 - \lambda_4 \end{array} \quad (30)$$

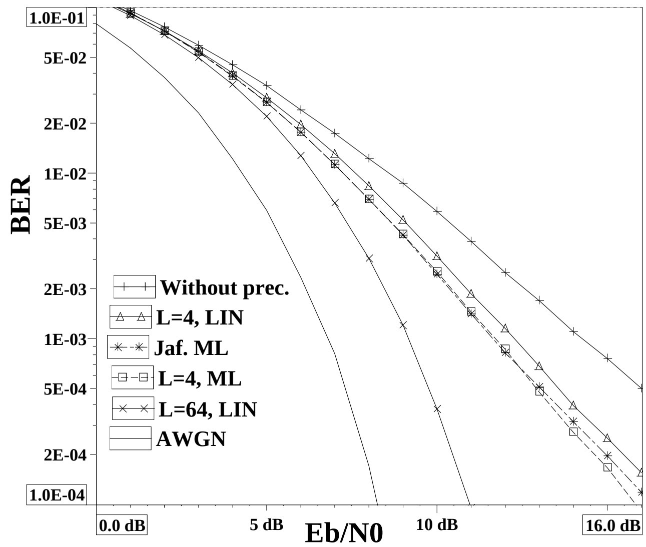

Fig. 2. Performance, without channel coding, of OSTBC with Hadamard linear precoding for η=2bps/Hz and 4-transmit antennas

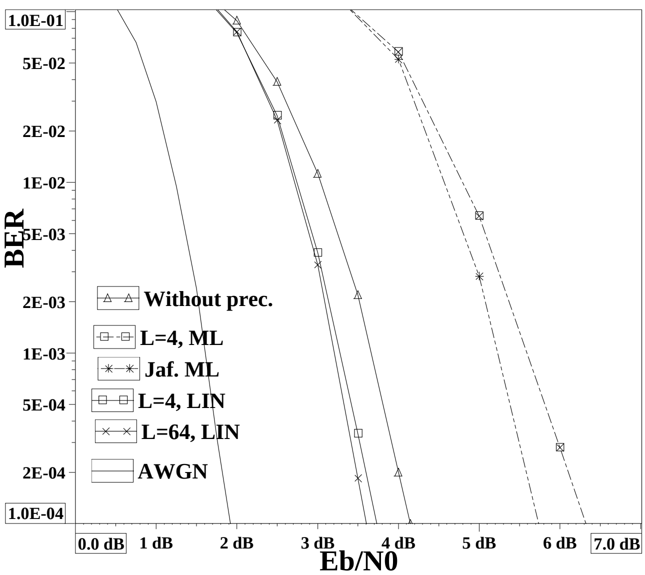

Fig. 3. Performance, with duo-binary turbo channel coding, of OSTBC with Hadamard linear precoding for η=2bps/Hz and 4-transmit antennas

We can note that $a+b = \sum_{i=1}^{8} |h_i|^2$ leading to a chi-square law with 16 degrees of freedom with interleaving compared to a chi-square law with 8 degrees of freedom without interleaving. $a-b = \sum_{i=1}^{4} |h_i|^2 - \sum_{i=5}^{8} |h_i|^2$ is one of the interference terms leading to a difference between chi-square laws with 8 degrees of freedom, and the others components follow the same law.

V. BER Simulation results

We carried out simulations over flat independent Rayleigh channels. Depending on the choice of $\mathbf{\Theta}_2$ matrix, simulations have demonstrated that the best performance are obtained when interference terms are either pure real or pure imaginary. This conclusion leads to choice for instance $\eta = \frac{\Pi}{4}$, $\theta_1 = \frac{5\Pi}{4}$ and $\theta_2 = \frac{3\Pi}{4}$. In addition, a Minimum Mean Square Error (MMSE) equalizer was demonstrated to provide the best performance for OSTBC decoding with linear precoding.

Applying the MMSE equalization, we get:

$$\lambda = \frac{\sum_{i=1}^{N_t} |h_i|^2}{\sum_{i=1}^{N_t} |h_i|^2 + \frac{1}{\gamma}} \tag{31}$$

where γ is the Signal to Noise Ratio at the receive antenna.

In Figures 2 and 3, we present BER performance obtained respectively with or without channel coding. The simulated systems implement $N_t = 4$ transmit and $N_r = 1$ receive antennas, excepted for the reference 2×1 Alamouti curve.

In these figures are represented several curves:

- "without prec" corresponds to the reference Alamouti OSTBC curve, applied without precoding technique for $N_t = 2$ and $N_r = 1$;
- "Jaf. ML" corresponds to the QOSTBC system implementing Jafarkhani codes and needing a ML detector at the reception side;
- "$L = 4$ ML" corresponds to our proposed scheme with a ML decoder using precoding Hadamard matrix of size 4;
- "$L = 4$ LIN" corresponds to our proposed scheme implementing a Linear decoder with Hadamard precoding matrix of size 4;
- "$L = 64$ LIN" corresponds to our proposed scheme implementing a Linear decoder with Hadamard precoding matrix of size 64;
- "AWGN" is the curve obtained with only Additive White Gaussian Noise.

Without channel coding, we observe that the specified precoded system with $L = 4$ outperforms of 2 dB the sole Alamouti scheme without precoding at $BER = 10^{-3}$. In fact, all diagonal terms $\frac{1}{2}\sum_{l=1}^{4} |h_l|^2$ follow a χ_8^2 chi-square law while the interference terms in $\frac{1}{2}(\sum_{l=1}^{2} |h_l|^2 - \sum_{l=3}^{4} |h_l|^2)$ follow a χ_4^2 difference. We can notice that for $L = 4$, the penalty when using linear detector instead of the ML detector is very small (around 0.5 dB at BER= 10^{-3}) but the complexity of the linear detector is quite lower than the ML detector. We remark that both ML detectors give quasi-equal performance. However, we observed that our proposed scheme outperforms all the other when L increases ($L = 64$). This result is due to the interference terms that decrease when L increases (cf. eq(20)) and that the diversity grows with L, following a χ_{2L}^2 law for diagonal terms. In fact, for $L = 64$, the slope of the curve is almost parallel to the gaussian curve.

With duo-binary turbo channel coding of half rate, we show that the performance of the turbo code associated with a ML space-time detector leads to worse performance than associated with a linear detector. We can explain this phenomena by the fact that at low Signal to Noise Ratio (SNR), the ML detector, that provides hard decisions to the turbo decoder, may also provides bad decisions to it; the latter being very sensitive to the input values gives in this turn false decisions. The systems with ML detector could be improved with the use of soft decisions or of fiability values but at a price of an additional complexity.

The best performance results are obtained when linear detectors are implemented. In fact, with linear precoding and orthogonal Almaouti STBC no hard decision is taken until duo binary channel decoding. At a low SNR, corresponding to a Bit Errror Rate of about 5.10^{-2} without channel coding, we notice that the performance of the linear detector systems are very close. That's why, the difference between performances after channel turbo-decoding are very small between the 3 linear detectors. Our proposed schemes whith linear precoding, ($L = 4$ or $L = 64$) have a $0.5dB$ performance gain compared to the simple 2×1 Alamouti system. Moreover, the use of a precoding matrix allows to achieve better performance even with a very low value of precoding matrix owing to the additional diversity brought to the turbo-decoding. We can see that the conclusion about performance results obtained with or without channel coding can be different. In fact, when analysing results without channel coding, we have to compare the different techniques by taking into account the activing area of the channel decoder and to exploite efficiently all availble diversities at the lowest price for complextity. The linear precoding system can also be efficiently applied to other antenna systems (2 or more) offering a large choice of global system depending on the the channel characteristics.

VI. Conclusion

In this paper, we have proposed a new scheme relying on the combination of OSTBC, issued on the Alamouti scheme, and a linear precoding, based on Hadamard construction matrices. This new scheme is applied to a 4 transmit antenna system and leads to an efficient exploitation of space-time diversity. In addition, as a linear detector is used at the reception side, the complexity linearly increases with the precoding matrix size but not exponentially compared to ML detectors often used in such precoding systems. Like this, our system described with 4-transmit antennas may be applied to other OSTBC codes and several antenna configurations. The performance results show that a simple linear detector is sufficient when using an OSTBC combined with linear precoding. This conclusion remains valuable since a turbo channel coding is used. In addition, we have seen that the conclusions about systems can be different when channel coding is applied or not, depending to the activing channel detector area, i.e. at a BER of about 5.10^{-2} before decoding. Our proposed scheme can be combined to other OSTBC and can easily be fitted into multi-carrier systems (OFDM or MC-CDMA). Therefore, these precoders can be applied to various MISO or MIMO transmissions in order to exploit spatial, temporal and frequency diversities. To conclude, we can say that the use of the Alamouti scheme for two transmit antennas or our proposed scheme for two or four transmit antennas, are good choices of transmission chain for future wireless communication systems.

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